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Induction Starter/ Alternators (ISAs) for Electric Hybrid Vehicles (EHVs)

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7.1 EHV Configuration

In this book, EHV stands for electric hybrid vehicles. EHV constitutes an aggressive novel technology aimed at improving comfort, gas mileage, and environmental performance of road vehicles [1,2].

The degree of “electrification” in a vehicle may be defined by the electric fraction, %E [3]:

$$\%E = \frac{\text{Peak electric power}}{\text{Peak electric power} + \text{Peak ICE power}} = \frac{P_{(el)}}{P_{(el)} + P_{(ICE)}} \quad (7.1)$$

For a mild hybrid car with battery soft-replenishing %E is lower than 40% in town driving. It may reach up to 70% when the battery is replenished from the power grid daily. %E becomes 100% for fully electric vehicles, with fuel cells or batteries or inertial batteries (flywheels) as the energy storage system.

The larger the electric fraction %E, the lower the internal combustion engine (ICE) rating (it is zero for a fully electric vehicle).

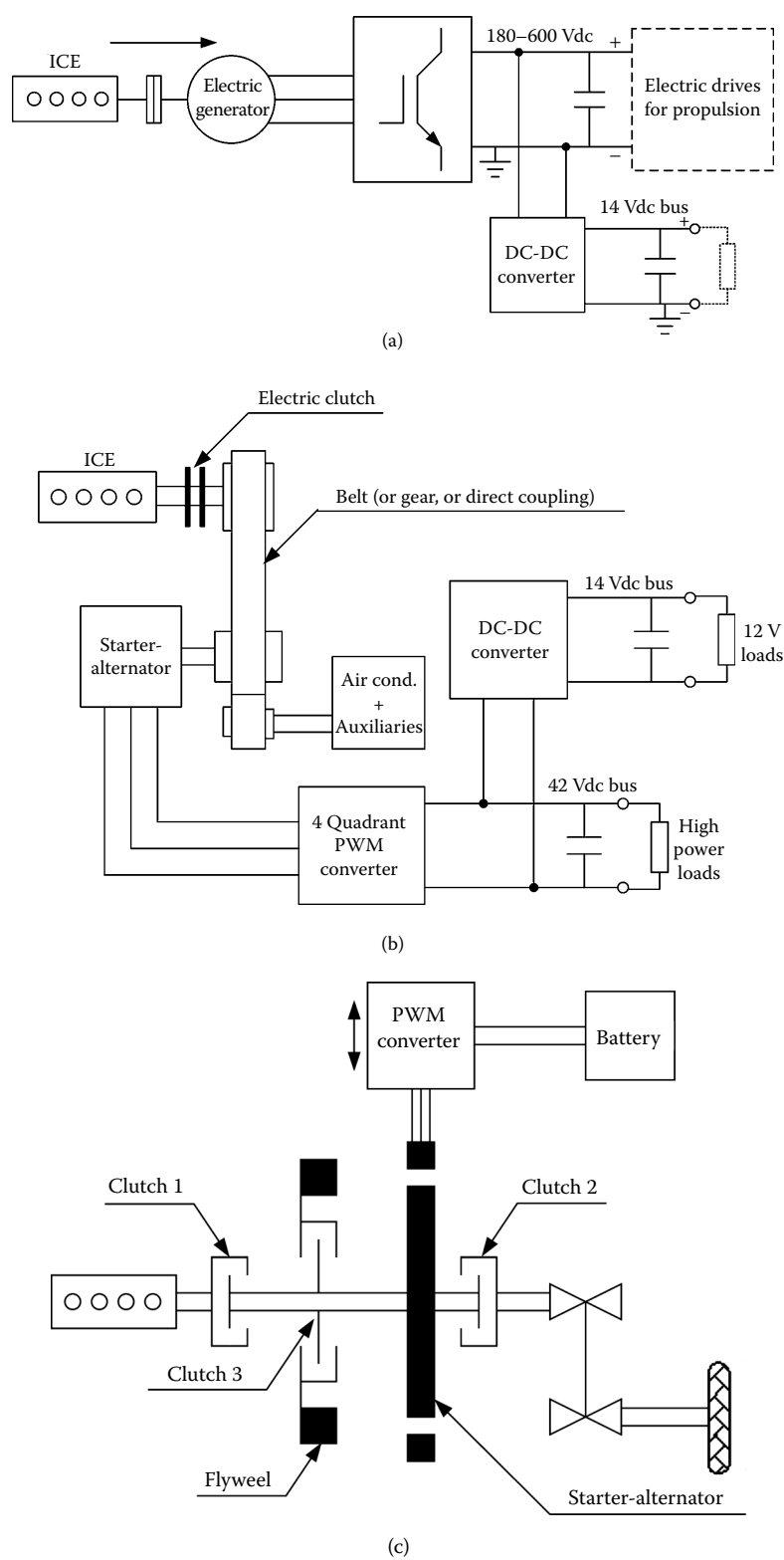


FIGURE 7.1 Basic vehicle electrification configurations: (a) series hybrids, (b) and (c) parallel hybrids, and (d) electric.

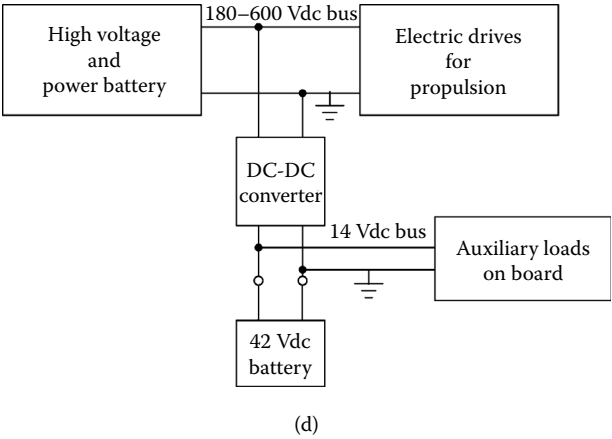


FIGURE 7.1 (Continued).

The electrification of vehicles is approached through a plethora of system configurations that may be broken down as illustrated in Figure 7.1a through Figure 7.1d.

In series hybrids (Figure 7.1a), the full-size ICE drives an electric generator on the vehicle that then produces electric energy for all tasks, from the electric drives to auxiliaries and battery recharging.

In parallel hybrids (Figure 7.1b and Figure 7.1c), the downsized ICE is started by the starter/alternator that then assists in propulsion at low to medium speeds and, respectively, works as a generator to feed the electrical loads and recharge the battery.

In fully electric vehicles (Figure 7.1d), a large high-voltage battery, recharged from the power grid once every day, supplies all electric drives used for vehicle propulsion. It also contains a 42 V_{dc} battery that supplies the auxiliaries. This latter battery is recharged from the main battery through a dedicated direct current (DC)–DC converter.

In mixed hybrids (Figure 7.1d), two or more motor/generators are used, for example, to electrically drive the front and the rear wheels (Figure 7.2) [3].

In all hybrid (and electric) vehicles, the air-conditioning and some auxiliaries should remain on duty during idle stop. Idle stop is stopping the engine or electric driving during halts in traffic jams or at traffic stoplights (Figure 7.1b).

Also, depending on the electric fraction %E and vehicle size (car, bus, and truck), the specifications vary within a large range.

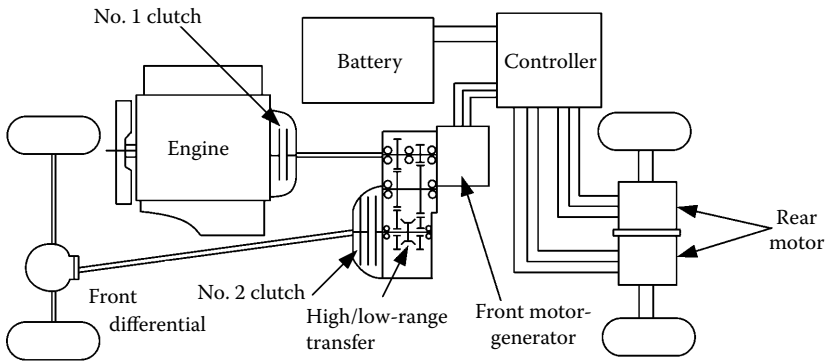


FIGURE 7.2 Mixed hybrid with front and rear motor/generator.

The placing of the electrical machine on the ICE shaft or its coupling with an additional transmission (belt or gear) is essential in the design of the starter/alternator, as the peak torque will depend on this transmission ratio $k_{el} \geq 1$.

The pulse-width modulator (PWM) converter peak kilovoltampere (kVA) rating depends heavily on the starter/alternator design, as the peak current required for peak torque “defines” the converter costs for given battery voltage.

Defining pertinent specifications and design optimization multiobjectives for the starter/alternator (motor/generator) system is of utmost importance.

7.2 Essential Specifications

Essential specifications for starter/alternators (motor/generator) on EHV are considered here to be the following:

- Starter/alternator functions
- ICE to starter/generator transmission ratio k_{el}
- Peak torque vs. speed for motoring
- Peak generator power (torque) vs. speed
- “Battery” voltage
- Battery self- or independent-replenishing method

7.2.1 Peak Torque (Motoring) and Power (Generating)

The peak torque for motoring is defined as the engine starting torque at 20°C and varies between 120 to 300 Nm for cars, but it may reach 1200 Nm for buses. This peak torque level has to be sustained up to $n_b = 250$ to 400 rpm for mild hybrids, and up to $n_b = 1000$ to 1200 rpm for full hybrids. Above base speed n_b , a constant peak power, up to maximum speed n_{max} , has to be provided:

$$P_{ekm} = T_{ekm} \cdot 2\pi \cdot n_b \quad (7.2)$$

The larger the n_{max}/n_b is, the larger the contribution of propulsion and its impact on fuel consumption reduction in city driving. This ratio n_{max}/n_b in motoring may range from 3:1 to 6:1. The larger the better for HEV performance, but this comes at the price of stator/alternator or PWM converter oversizing.

A large constant peak power range imposes a few design solutions for which the whole system — starter/alternator, battery, and power converter — costs, size, and losses all have to be simultaneously considered.

A multiratio mechanical transmission reduces the constant power speed range of the starter/alternator and allows for a smaller size (volume) electrical machine at the price of the additional costs for a more complex transmission.

Typical motoring torque/speed and generating peak power/speed requirements for a mild hybrid (42 V_{dc}, $I_{peak} < 170$ A) small car are shown in Figure 7.3 [4–9].

The limit of 170 A on the battery is based on the acceptable voltage drop (losses) in the 42 V_{dc} battery pack made of three standard lead acid car batteries in series.

The generating power limit is based on the battery receptivity and PWM-converter-controlled starter/generator limits to safely deliver power at high speed at limited battery overvoltage.

Without winding changeover (reconfiguration), a 3/1 constant power speed range is possible without machine or converter oversizing. A 5/1 constant power speed range ($n_{max}/n_b = 5/1$) is visible in Figure 7.4 and, thus, oversizing and winding changeover are required. Above 2500 rpm, in Figure 7.4, the driving power decreases due to “lack” of voltage in the battery to sustain it.

The specific propulsion requirements in a fully electric car with standard multispeed transmission are shown in Figure 7.4.

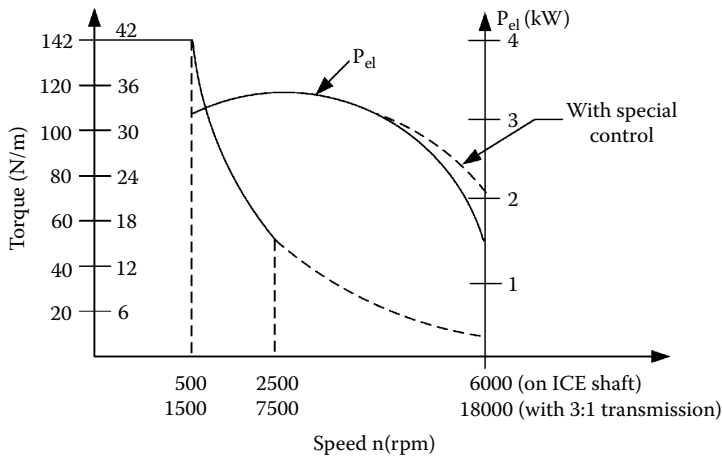


FIGURE 7.3 Potential mild hybrid car peak torque and speed motoring and power and speed generating envelopes.

For an electric city bus, 75 kW of electric propulsion is considered in [Figure 7.5](#). A single-stage 6.22 gear reduction ratio is mentioned in the literature [10].

7.2.2 Battery Parameters and Characteristics

The main battery parameters are as follows:

- The battery capacity: Q
- The discharge rate: Q/h
- The state of charge: SOC
- The state of discharge: SOD
- The depth of discharge: DOD

The amount of free electrical charge generated to the active battery material at the negative electrode ready to be consumed by the positive electrode is called the battery capacity Q .

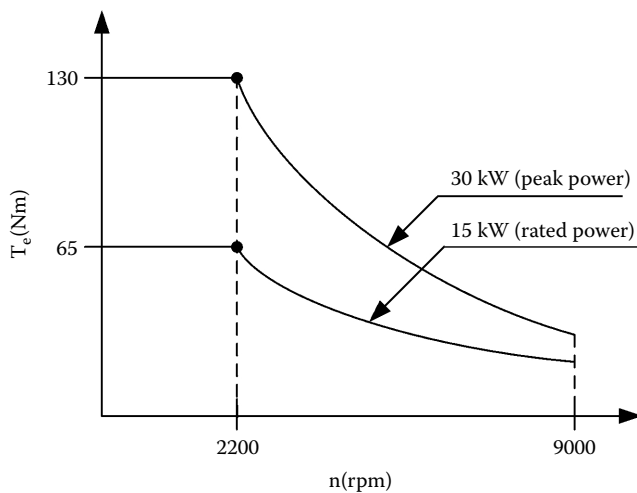


FIGURE 7.4 Typical motoring torque and speed envelopes for a fully electric car.

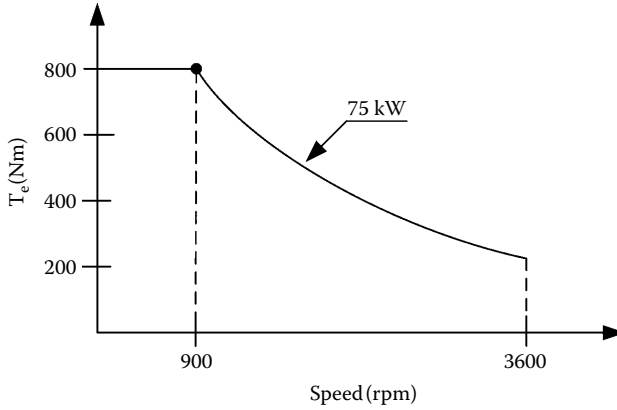


FIGURE 7.5 Typical rated torque and speed envelopes for a fully electric city bus.

Q is measured in amperehours (Ah); 1 Ah = 3600 C; 1 C is the charge transferred by 1 A in 1 sec. The theoretical capacity Q_T is as follows:

$$Q_T = x \cdot n \cdot F; \quad F = L \cdot e_0 \quad (7.3)$$

where

x is the number of moles of reactant for complete discharge

n is the number of electrons released by the negative electrode during discharge

$L = 6.022 \times 10^{23}$ is the number of atoms per mole (Avogadro's constant)

$e_0 = 1.601 \times 10^{-19} \text{ C}$ is the electron charge (F is the Faraday constant)

$$Q_T = 27.8 \cdot x \cdot n [\text{Ah}] \quad (7.4)$$

With a number of cells in series, the capacity of a cell equals the capacity of the battery.

The discharge current is called the discharge rate $Q[\text{Ah}]/h$, where h is the discharge rate in hours. If a 200 Ah battery discharges in half an hour, the discharge rate is 400 A.

The state of charge (SOC) represents the battery capacity at the present time:

$$\text{SOC}(t) = Q_T - \int_0^t i(\tau) d\tau \quad (7.5)$$

The state of discharge SOD(t) represents the charge already drawn from a fully charged battery:

$$\text{SOD}(t) = \int_0^t i(\tau) d\tau = Q_T - \text{SOC}(t) \quad (7.6)$$

The depth of discharge (DOD) is the per unit (P.U.) battery discharge:

$$\text{DOD}(t) = \frac{\text{SOD}(t)}{Q_T} = \frac{\int_0^t i(\tau) d\tau}{Q_T} \quad (7.7)$$

A deep discharge takes place when $\text{DOD} > 0.8$ (80%).

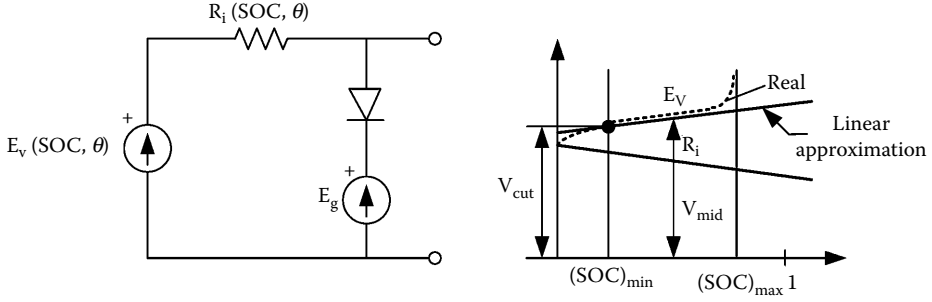


FIGURE 7.6 Simplified battery model.

Adequate modeling of the battery is essential in starter/alternator design, as the battery voltage varies with temperature, SOC (SOD) and its recharging process set limits on the electric energy recovered in the generating mode.

The simplest battery model contains an electromagnetic field (emf), E_v and an internal resistance R_i are both dependent on battery SOC and on temperature θ (Figure 7.6) [11].

The battery emf E_v increases with the SOC (and decreases with the SOD), while the internal resistance does the opposite.

When the battery is deeply discharged (SOC_{min}), the voltage tends to drop steeply. This is the cut voltage V_{cut} beyond which the battery should not, in general, be used. The practical capacity is thus,

$$Q_p = \int_0^{t_{\text{cut}}} i(t) dt < Q_T \quad (7.8)$$

When the battery is started, the constant discharge current should also be specified.

The emf E_v decreases when the temperature increases. There is also a step increase in E_v when the SOC is high (Figure 7.6). This partially explains, for example, the denomination of 14 (42) V_{dc} batteries when, on load, they work as 12 (36) V_{dc} .

The electrical energy extracted from the battery W_b is as follows:

$$W_b = \int_0^{t_{\text{cut}}} V \cdot I dt \quad (7.9)$$

With constant discharging current, the total battery voltage vs. time looks as shown in Figure 7.7. The discharge time to V_{cut} is larger if the discharge current is smaller.

For constant discharge current (I) the cut-time t_{cut} is offered by Peukert's equation [2]:

$$t_{\text{cut}} = \frac{C_c}{I^{n_c}} \quad (7.10)$$

with C_c and n_c as constants.

The specific energy (Wh/kg) is the discharge energy W_b per battery weight. For lead acid batteries, the specific energy is around 50 Wh/kg at $Q/3$ rate.

Other batteries have higher energy densities, but their costs per watthour tend to be higher, in general.

The battery power $P(t)$ is as follows:

$$P(t) = V_t \cdot i = (E_v - R_i i) \cdot i \quad (7.11)$$

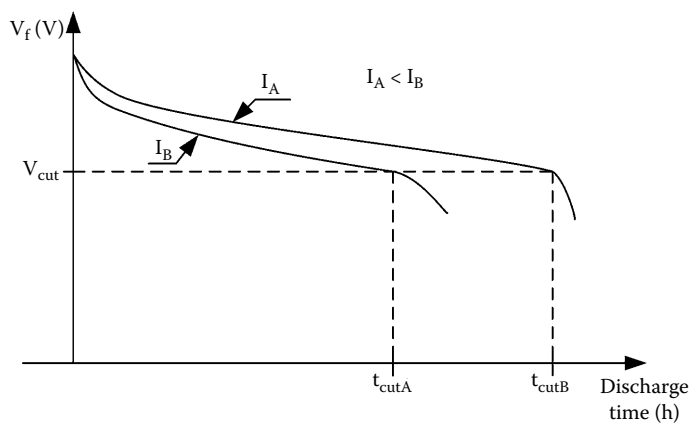


FIGURE 7.7 Voltage per time for constant current discharge.

For constant E_v , the maximum power P_{tmax} occurs, as known, for $R_{load} = R_i$:

$$P_{tmax} = \frac{E_v^2}{4 \cdot R_i} \tag{7.12}$$

The rated instantaneous power P_{ti} is the maximum power deliverable for a short discharge time without damage, while instantaneous power P_{ic} corresponds to large discharge intervals and no damage to the battery.

The specific power is P_i/M_B (W/Kg). The lead acid battery may deliver a maximum of 280 to 400 W/kg at DOD = 80%. Other batteries may produce less.

Table 7.1 presents typical characteristics of a few batteries for EHV's.

Note that fuel cells, inertial (flywheels) batteries, and supercapacitors may act as alternative energy storage systems on vehicles. They are characterized by smaller energy density but higher power density (2 kW/kg) [12]. The fuel cells tend to have smaller efficiency (60 to 70%), while inertial and supercapacitor batteries have higher efficiency.

Super-high-speed inertial (flywheel) batteries, in vacuum, with (eventually) magnetic bearing, should surpass the batteries in all ways, including the possession of a long life and a 2 to 3 min recharge time. Inertial batteries contain a super-high-speed generator/motor for recharging, controlled through a bi-directional PWM static power converter. This subject will be treated in the chapter on super-high-speed generators (Chapter 10).

TABLE 7.1 Typical Characteristics of a Few Batteries for Electric Hybrid Vehicles (EHV)					
Battery	Wh/kg	W/kg	Efficiency %	Cycle life	Cost \$/kWh
Lead acid	35–50	150–400	80	500–1000	100–150
Nickel–cadmium	30–50	100–150	75	1000–2000	250–350
Nickel–metal–hydride	60–80	200–300	70	1000–2000	200–350
Aluminum–air	200–300	100	50	—	—
Zinc–air	100–220	30–80	60	500	90–120
Sodium–sulfur	150–240	240	85	1000	200–350
Sodium–nickel–chloride	90–120	140–160	80	1000	250–350
Lithium–polymer	150–200	350	—	1000	150
Lithium–ion	80–130	200–300	95	1000	200

Source: Adapted from I. Husain, *Electric and Hybrid Vehicles*, CRC Press, Boca Raton, FL, 2003.

In view of the many EHV schemes and ratings, in the following paragraphs, we will concentrate on various aspects of induction machine modeling, and design for variable speed and control for starter/generator applications. In other words, we will aim to provide a comprehensive tool for EHV designers.

A numerical example emphasizes the essentials and gives a feeling of the magnitude of this topic.

7.3 Topology Aspects of Induction Starter/Alternator (ISA)

The operations of ISA are characterized by the following:

- A thermally harsh environment
- Limited volume (weight)
- Requirements for high efficiency and power factor
- Maximum speed above 6000 rpm
- Bidirectional converter supply to and from a DC bus voltage source (battery); battery voltages vary by about (or more than) 30%, depending on SOC, load, and ambient temperature; the DC voltage goes up with power rating from 42 V_{dc} to 600 V_{dc}

These operating conditions lead to some specifications in ISA design and control.

First, squirrel-cage rotors are to be used. Also, the high speed corroborated with low volume leads to a large number of poles. As the maximum speed goes up, the number of poles should go down to limit the maximum fundamental frequency f_1 to 500 to 600 Hz. The frequency limitation is prompted by both reasonable iron losses and PWM converter switching losses and costs.

The number of poles is larger when the rotor diameter (and peak torque) is larger, as what counts is the ratio of the pole pitch τ to airgap g , to provide a reasonable magnetization current (power factor).

In general, $2p_1 = 8$ to 12 for $n_{max} < 6000$ rpm and $2p_1 = 4$ to 6 for $n_{max} > 12,000$ rpm.

Higher than 6000 rpm speeds are typical when the belt or gear transmission with $k_e = 2/1$ to 3/1 is used to couple the ISA to the ICE.

Maximum fundamental frequencies of $f_1 = 500$ to 600 Hz lead to rotor frequencies of $f_2 = 5$ to 6 Hz, even for a slip $S = 0.01$ ($f_2 = S \cdot f_1$).

Skin effect has to be limited both in the stator and in the rotor because of the large fundamental maximum frequencies and due to higher current time harmonics incumbent to PWM converter supplies. Also, to keep the losses down, while very large torque densities are required, the rotor resistance has to be reduced by design.

The stator slots should be semiclosed and rectangular or trapezoidal. Rotor slot shapes with low skin effect should be chosen (Figure 7.8a through 7.8d).

The U-bridge closed rotor slot (Figure 7.8d) is supposed to reduce the surface and flux pulsation losses in the outer part of the rotor cage. Unfortunately, this merit is counteracted by a larger slot leakage inductance. Finally, breakdown torque is reduced this way.

Using copper instead of aluminum may help in reducing the cage losses, despite the fact that the skin effect tends to increase due to higher conductivity of copper in comparison with aluminum. Also, smaller cross-sectional rotor bars would allow more room for rotor teeth, leading to reduced rotor core saturation. Insulating the copper bars from slot walls may also prove useful in reducing the interbar rotor current losses.

With $2p_1 = 8$ to 12 poles, in most cases, the number of slots per pole and phase $q_1 = 2, 3$. Only for $2p_1 = 4$ to 6 poles, does $q_1 = 3$ to 5.

Chorded coils in the stator are required to reduce the fifth and seventh magnetomotive force (mmf) space harmonics with their rotor core surfaces and cage losses.

Skewing is also an option in reducing the first slot harmonics $v = 6q \pm 1$ with their rotor core surface and pulsation additional losses. When the number of rotor slots N_r is chosen to be smaller than the number of stator slots N_s ($N_r < N_s$):

$$0.8 < \frac{N_r}{N_s} < 1 \quad (7.13)$$

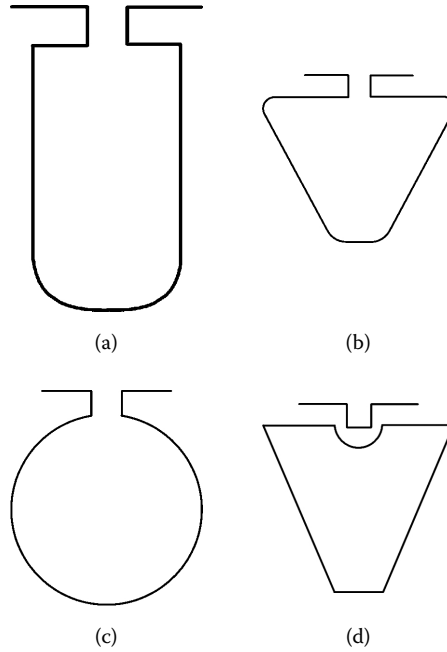


FIGURE 7.8 Suggested induction starter/alternator (ISA) rotor slots: (a) rectangular, (b) trapezoidal, (c) circular, and (d) U bridge closed slot.

and N_s , N_r restrictions observed for reducing the parasitic torque harmonics and noise (Reference [13], chap. 11), no skewing is required, even if the rotor cage is not isolated from the rotor core.

To reduce the end-ring leakage permeance, the end-rings should be placed at a distance with respect to the core.

When the ISA is placed on the engine shaft, the rotor diameter D_{in} should be higher than a given value, as the respective space is required for cooling.

The environment in such a direct coupling has an ambient temperature $T_{amb} \approx 110$ to 130°C . So, even with Class $F(H)$ insulation material, the winding overtemperature $\Delta\theta_{amb} \leq 60 - 70^\circ\text{C}$. The rotor temperature, for prolonged full generating, may reach up to 240°C . Forced cooling may be needed to fulfill such conditions, especially with the ISA designed for low volume at the peak torque.

The peak current density in the stator conductors may reach values around or even larger than 30 A/mm^2 .

Further on, the degree of magnetic saturation in ISA, for peak torque, should be high for the same reason. A peak airgap flux density fundamental of 1.0 to 1.15 T is common for ISA.

Allowing uniform, though advanced, magnetic saturation in the stator and rotor teeth and yokes seems to be the key to small airgap space harmonics in the airgap flux density (Reference [13], p. 125). Small space harmonics in the airgap flux density lead to small such harmonics in the teeth and yokes. So, in fact, the iron core harmonics losses due to space harmonics are reduced this way, and so are the torque pulsations, vibration, and noise.

In terms of analysis, the gain is exceptional. The phasor and space-phasor ($d-q$) model could be used by simply making use of the magnetization curve of the machine, calculated (and measured) on no-load to adopt the machine's main field inductance.

For the peak current, to produce limited slot-leakage tangential tooth-stop flux density B_{tt} (Figure 7.9a through Figure 7.9c), the slot opening W_{os} is kept at 5 to 6g (g is the airgap).

Also, for open and semiopen slots, the slot opening W_{os} should be "reduced" to W'_{os} , making use of a magnetic wedge with a relative permeability $\mu_{rw} = 2 \div 4$, $W'_{os} = W_{os} / \mu_{rw}$.

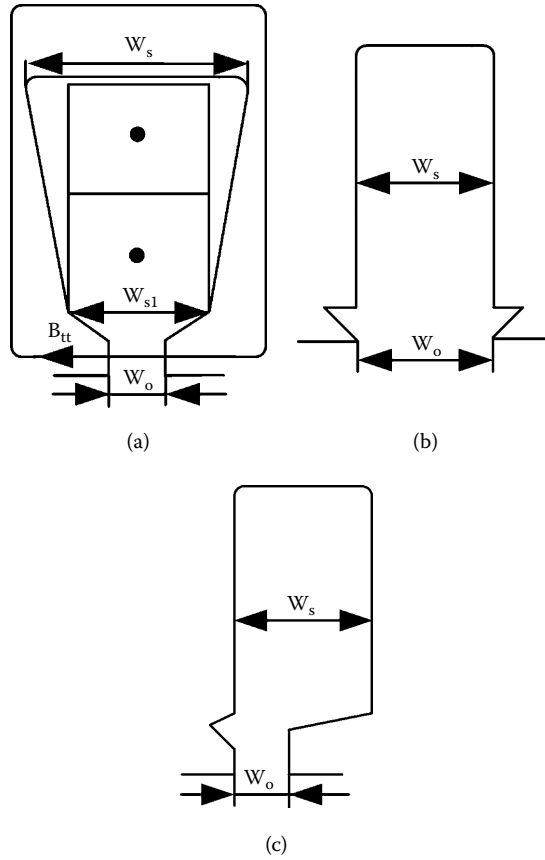


FIGURE 7.9 Typical stator slots for induction starter/alternator (ISA): (a) semiclosed trapezoidal, (b) open rectangular, and (c) semiopen rectangular.

Allowing for $W_{os}'/g > 6$ would imply a notably larger magnetization current and larger teeth flux pulsation that induces large rotor surface and harmonics cage losses.

Open slots allow for the machine placing of coils in slots and allow for cuts to be made in manufacturing costs and time. The optimum airgap is a trade-off between magnetization current and rotor surface harmonics iron losses. In general, for the peak torque in the range from 40 to 1200 Nm, the airgap may vary between 0.5 and 1.0 mm.

7.4 ISA Space-Phasor Model and Characteristics

As already pointed out in the previous paragraph, uniform and heavy magnetic saturation (by design) of stator and rotor teeth and yokes leads to a close to sinusoidal airgap flux distribution.

Consequently, the space-phasor model, so typically used in modeling induction motor drives [14], may be applied even for calculating torque, provided the magnetization curve,

$$\bar{\Psi}_m(I_m) = L_m(I_m) \cdot \bar{I}_m \quad (7.14)$$

is known either from calculations or from tests.

$\bar{\Psi}_m$ is the airgap flux linkage space phasor, and $\bar{I}_m$ is the airgap-flux magnetization space-phasor current:

$$\bar{I}_m = \bar{I}_s + \bar{I}_r \quad (7.15)$$

$\bar{I}_s$ is the stator space-phasor current (stator current vector), and $\bar{I}_r$ is the rotor current vector:

$$\bar{I}_s^b = \frac{2}{3} \cdot \left(i_a + i_b \cdot e^{j\frac{2\pi}{3}} + i_c \cdot e^{-j\frac{2\pi}{3}} \right) \cdot e^{-j\theta_s^b} \quad (7.16)$$

Equation 7.16 represents the so-called Park complex transformation, with θ_s^b equal to the position angle of the orthogonal axis of coordinate system with respect to phase a axis in the stator. The same transformation is valid for the stator voltage and flux vectors.

The stator and rotor flux $\bar{\Psi}_s, \bar{\Psi}_r$ are, simply,

$$\begin{aligned} \bar{\Psi}_s &= L_{sl} \cdot \bar{I}_s + \bar{\Psi}_m \\ \bar{\Psi}_r &= L_{rl} \cdot \bar{I}_r + \bar{\Psi}_m \end{aligned} \quad (7.17)$$

L_{sl} and L_{rl} are the leakage inductances, both reduced to the stator, and so are $\bar{I}_r$ and $\bar{\Psi}_r$.

The stator voltage vector equation in stator coordinates is as follows:

$$\bar{I}_s \cdot R_s - \bar{V}_s = -\frac{d\bar{\Psi}_s}{dt} \quad (7.18)$$

And for the rotor, in rotor coordinates,

$$\bar{I}_r \cdot R_r = -\frac{d\bar{\Psi}_r}{dt} \quad (7.19)$$

For general coordinates,

$$\begin{aligned} \bar{I}_r^b &= \bar{I}_r^r \cdot e^{-j(\theta_s^b - \theta_{er})} & \bar{\Psi}_r^b &= \bar{\Psi}_r^r \cdot e^{-j(\theta_s^b - \theta_{er})} \\ \bar{I}_s^b &= \bar{I}_s^s \cdot e^{-j\theta_s^b} & \bar{\Psi}_s^b &= \bar{\Psi}_s^s \cdot e^{-j\theta_s^b} \\ \bar{V}_s^b &= \bar{V}_s^s \cdot e^{-j\theta_s^b} \end{aligned} \quad (7.20)$$

θ_{er} is the rotor position with respect to stator phase a in electrical degrees ($\theta_{er} = p_1 \cdot \theta_r$).

Equation 7.18 and Equation 7.19 thus become

$$\begin{aligned} \bar{I}_s \cdot R_s - \bar{V}_s &= -\frac{\partial \bar{\Psi}_s}{\partial t} - j \cdot \omega_b \cdot \bar{\Psi}_s & \omega_b &= d\theta_s^b / dt \\ \bar{I}_r \cdot R_r &= -\frac{\partial \bar{\Psi}_r}{\partial t} - j \cdot (\omega_b - \omega_r) \cdot \bar{\Psi}_r & \omega_r &= d\theta_{er} / dt = p_1 \cdot \omega_r \end{aligned} \quad (7.21)$$

The superscript b has been dropped in Equation 7.21, which are now both written in general coordinates that rotate at speed ω_b .

For steady-state stator voltages $V_{a,b,c}$,

$$V_{a,b,c} = V_1 \cdot \sqrt{2} \cdot \cos\left(\omega_1 t - (i-1) \cdot \frac{2\pi}{3}\right) \quad (7.22)$$

After applying the Park complex transformation of Equation 7.16, we obtain

$$\bar{V}_s = V_1 \cdot \sqrt{2} \cdot \cos(\omega_1 - \omega_b) \cdot t - j \cdot V_1 \cdot \sqrt{2} \cdot \sin(\omega_1 - \omega_b) \cdot t \quad (7.23)$$

So, the frequency of the voltage vector applied to the induction machine at steady state $(\omega_1 - \omega_b)$ depends on the speed of the orthogonal reference ω_b .

Three reference system speeds are most used:

- Stator coordinates: $\omega_b = 0$
- Rotor coordinates: $\omega_b = \omega_r$
- Synchronous coordinates: $\omega_b = \omega_1$

Synchronous coordinates are used for machine control simulation for vector (flux-oriented) control (FOC) implementation while stator coordinates are used for direct torque and flux control (DTFC). All three values of ω_b are used for building state observers for FOC or direct torque and flux control (DTFC).

For synchronous coordinates, steady state means zero frequency in Equation 7.23, as $\omega_b = \omega_1$.

Constant rotor flux in Equation 7.21 means $\partial \Psi_r / \partial t = 0$ and constitutes the basis for vector control. Let us consider synchronous coordinates: $\omega_b = \omega_1$. This means that only the amplitude of the rotor flux vector has to be maintained constant to eliminate rotor electrical transients. It was proven that constant rotor flux control leads to the fastest torque transients in the machine fed through a controlled current source. DTFC does almost the same but in stator coordinates.

Eliminate the stator flux and rotor current in Equation 7.21 by making use of Equation 7.15 and Equation 7.17, and replace $\partial/\partial t$ with s (Laplace operator):

$$\begin{aligned} \bar{I}_s \cdot (R_s + (j \cdot \omega_1 + s) \cdot L_{sc}) - \bar{V}_s &= - (s + j \cdot \omega_1) \cdot \frac{L_m}{L_r} \cdot \bar{\Psi}_r; \\ L_{sc} &= L_{sl} + L_m \\ \bar{I}_s \cdot R_r &= (R_r + (s + j \cdot S \cdot \omega_1) \cdot L_r) \cdot \frac{\bar{\Psi}_r}{L_m}; \\ L_r &= L_{rl} + L_m \\ L_{sc} &= L_s - \frac{L_m^2}{L_r} \\ S &= (\omega_1 - \omega_r) / \omega_1 \text{ -- the slip} \end{aligned} \quad (7.24)$$

The torque is

$$T_e = \frac{3}{2} \cdot p_1 \cdot \text{Real}(j \cdot \bar{\Psi}_s \cdot \bar{I}_s^*) = \frac{3}{2} \cdot p_1 \cdot L_m \cdot \text{Real}(j \cdot \bar{I}_s \cdot \bar{I}_r^*)$$

For constant rotor flux, $s = 0$ in the second expression of Equation 7.24. Also, for steady state $s = 0$ in the first equation, synchronous coordinates

$$\bar{I}_{so} \cdot (R_s + j \cdot \omega_1 \cdot L_{sc}) - \bar{V}_{so} = -j \cdot \omega_1 \cdot \frac{L_m}{L_r} \cdot \bar{\Psi}_{ro} \quad (7.25)$$

$$\bar{I}_{so} = \frac{\bar{\Psi}_r}{L_m} + j \cdot S \cdot \omega_1 \cdot T_r \cdot \frac{\bar{\Psi}_r}{L_m} = I_M + j \cdot I_T; \quad (7.26)$$

$$T_r = \frac{L_r}{R_r} \quad (7.27)$$

For constant rotor flux, the stator flux vector $\bar{\Psi}_s$ is as follows:

$$\bar{\Psi}_s = L_s \cdot I_M + j \cdot L_{sc} \cdot I_T \quad \bar{\Psi}_r = L_m \cdot I_M \quad (7.28)$$

and the torque T_e is

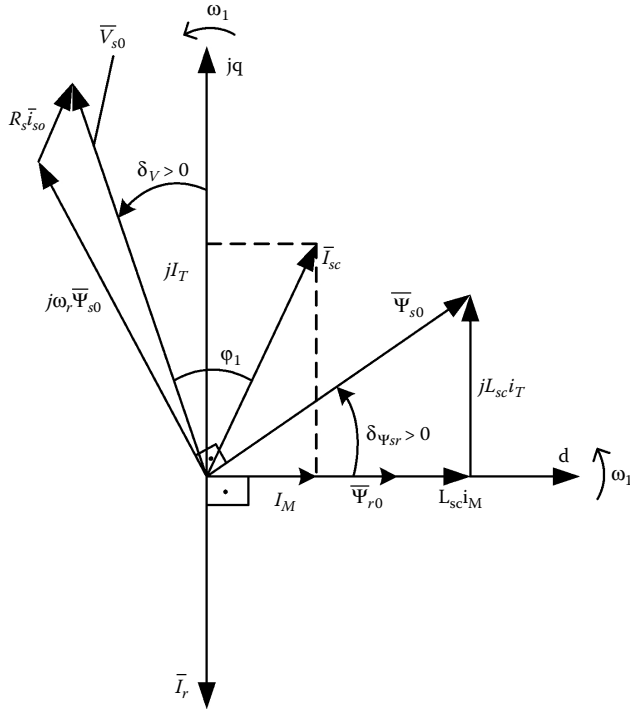
$$T_e = \frac{3}{2} \cdot p_1 \cdot (L_s - L_{sc}) \cdot I_M \cdot I_T; \quad I_r = -j \frac{S \cdot \omega_1 \cdot \bar{\Psi}_r}{R_r} \quad (7.29)$$

A few remarks are in order after this theoretical marathon:

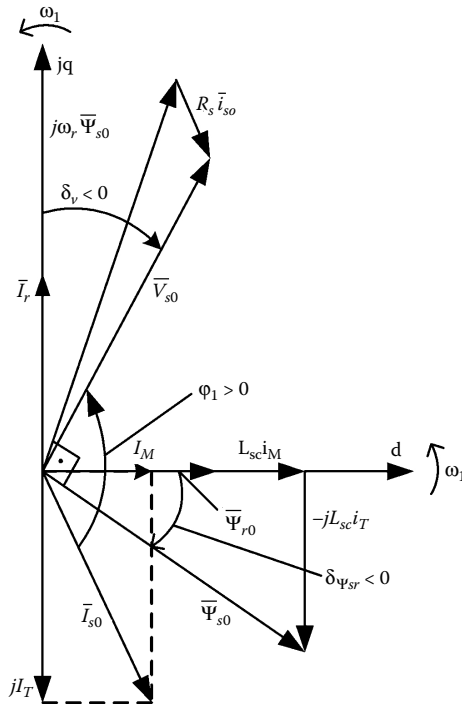
- The torque expression of the induction machine for constant rotor flux (amplitude only in synchronous coordinates) resembles that of a synchronous reluctance machine, where the d axis inductance is the no-load inductance L_s , and the q axis inductance is the short-circuit inductance L_{sc} . The higher the difference $L_s - L_{sc}$, the larger the torque for given current.
- For constant rotor flux, the stator current vector has two components: a flux one I_M aligned with the rotor flux vector and the torque one I_T , 90° ahead in the direction of motion for motoring ($S > 0$) and behind ($S < 0$) for generating.
- For constant rotor flux, the stator flux vector has two components (Equation 7.28) produced by the I_M and I_T stator current components.

Equation 7.25 through Equation 7.29 may be represented in a vector diagram as in [Figure 7.10a](#) and [Figure 7.10b](#):

- For motoring (and braking) — $S > 0$ — the stator current vector $\bar{I}_{so}$ is ahead of stator flux vector $\bar{\Psi}_{so}$ in the direction of motion. The opposite is true for generating.
- Voltage, current, and flux vectors power angles $\delta_v, \delta_i, \delta_{\Psi}$ may be defined. For motoring, the stator flux vector $\bar{\Psi}_{so}$ is ahead of the rotor flux vector $\bar{\Psi}_{ro}$ along the direction of motion, and it is lagging for generating.
- The core losses are not yet considered, but for a first approximation, they depend only on $\omega_1, S\omega_1$ and $\bar{\Psi}_s$ or $\bar{\Psi}_m$. At low speed, the core losses are small, as ω_1 is small, though full flux is injected in the machine. At high speed, the frequency ω_1 is high, but the flux $\bar{\Psi}_s$ is small due to voltage limitation. When the winding or pole number changeover occurs at the changeover speed, the core losses are suddenly increased, as the flux $\bar{\Psi}_s$ is brought to the highest level, again limited only by magnetic saturation. The stator flux $\bar{\Psi}_s$ surpasses the airgap flux $\bar{\Psi}_m$ only by a few percent, in general (due to stator leakage flux $L_{sl}\bar{I}_s$).



(a)



(b)

FIGURE 7.10 Induction starter/alternator (ISA) vector diagram (steady state in synchronous coordinates): (a) motoring and (b) generating.

- The airgap flux Ψ_m magnetization current $\bar{I}_m$ is as follows:

$$\bar{I}_m = \bar{I}_s + \bar{I}_r = I_M + j \cdot I_T \cdot \left(1 - \frac{L_m}{L_r}\right) = \frac{\Psi_r}{L_m} \cdot \left(1 + j \cdot \frac{S \cdot \omega_1 \cdot L_{rl}}{R_r}\right) \quad (7.30)$$

$$\bar{\Psi}_m = \Psi_r - L_{rl} \cdot \bar{I}_r = \Psi_r \cdot \left(1 + j \cdot \frac{S \cdot \omega_1 \cdot L_{rl}}{R_r}\right) \quad (7.31)$$

It should be noticed that once the rotor flux Ψ_r and the torque values are set, the amplitude of the stator current I_s the slip frequency value $S\omega_1$, the airgap flux Ψ_m , and the stator flux Ψ_s amplitudes are all assigned certain values, depending also on the machine parameters $R_r, L_m(\Psi_m)$ and L_{rl}, L_{sl} .

- The magnetization inductance L_m depends on the airgap flux only — $L_m(\Psi_m)$ or $L_m(I_m)$ — while rotor temperature changes the rotor resistance R_r . The stator and rotor leakage inductances are considered constant, in general.
- At low speed, for peak torque, as the core losses are small, the maximum torque per current principle should be applied:

$$\begin{aligned} I_s &= \sqrt{I_M^2 + I_T^2} \\ T_e &\approx \frac{3}{2} \cdot p_1 \cdot (L_m - L_{rl}) \cdot I_M \cdot I_T \\ \frac{\partial T_e}{\partial I_M} &= 0 \end{aligned} \quad (7.32)$$

Only if $L_m = \text{constant}$, albeit heavily saturated,

$$I_M = I_T = \frac{I_{s_{ki}}}{\sqrt{2}} \quad (7.33)$$

The peak torque for given current is, thus,

$$\begin{aligned} T_{e_{ki}} &= \frac{3}{2} \cdot p_1 \cdot (L_m - L_{rl}) \cdot \frac{I_{s_{ki}}^2}{2} \\ (S \cdot \omega_1)_{ki} \cdot \frac{L_r}{R_r} &= 1 \\ \bar{I}_{m_{ki}} &= \frac{I_{s_{ki}}}{\sqrt{2}} \cdot \left(1 + j \cdot \frac{L_{rl}}{L_r}\right) \end{aligned} \quad (7.34)$$

The copper losses for the peak torque $T_{e_{ki}}, p_{Co_{ki}}$, are as follows:

$$p_{Co_{ki}} = \frac{3}{2} \cdot I_{s_{ki}}^2 \cdot \left(R_s + R_r \cdot \left(\frac{L_m}{L_r} \right)^2 \cdot \frac{1}{2} \right) \quad (7.35)$$

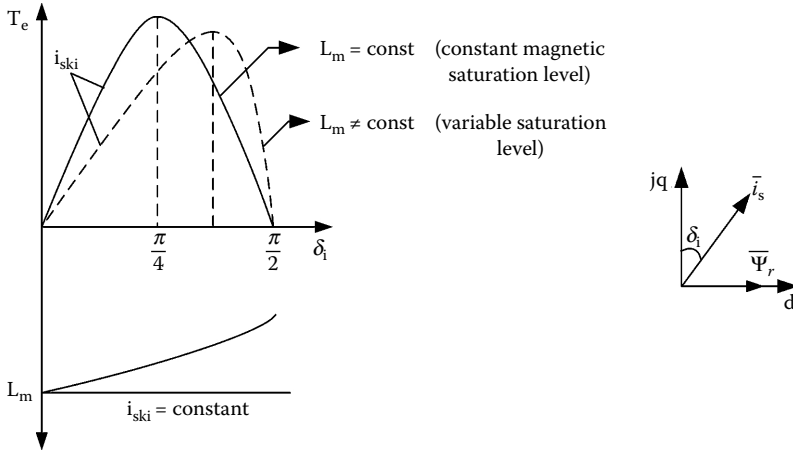


FIGURE 7.11 Torque vs. current power angle δ_i with constant and variable saturation.

It is now evident that L_m (and L_r) depend predominantly on the magnetization current of the main flux, $I_{m_{ki}}$.

As the short-time-lived peak torque should be large, so too will be the peak current. Consequently,

$$I_{m_{ki}} = \frac{I_{s_{ki}}}{\sqrt{2}} \quad (1.02 \text{ to } 1.1) \text{ should be very large.}$$

Current densities up to 30 to 40 A/mm² are admitted for peak torque short durations and adequate cooling.

Heavy oversaturation of the machine is thus mandatory, for very large peak torque, though L_m ($I_{m_{ki}}$) decreases due to heavy saturation.

An optimum saturation level for given geometry and torque (for given volume) is to be obtained.

The ideal current power angle $\delta_i = 45^\circ$ but, due to variable saturation, it departs from 45° to notably larger values, as L_m decreases with saturation to obtain maximum torque for given stator current (Figure 7.11).

The magnetic saturation curve may be calculated or measured and then curve fitted. A standard approximation is as follows [15]:

$$\Psi_m = \frac{L_o \cdot I_m}{1 + I_m / I_{mo}} + L_{oo} \cdot I_m = L_m \cdot I_m \quad (7.36)$$

So,

$$L_m(I_m) = \frac{L_o}{1 + I_m / I_{mo}} + L_{oo} \quad (7.37)$$

The constants L_o, L_{oo}, I_{mo} are to be determined to best fit the analytical or finite element field model of the saturated machine over the entire magnetization current range.

- As speed goes up, the machine cannot keep the peak torque past a certain speed, called the base speed ω_{rb} , for which full output voltage of the PWM converter is reached. In general, full flux is considered for base speed, full voltage, and peak torque. If the flux is reduced, the base speed may be increased for the cost of larger current. For $I_{M_{ki}} = I_{s_{ki}} / \sqrt{2} = I_{T_{ki}}$ the stator flux $\Psi_{s_{ki}}$ is as

follows:

$$\Psi_{s_{ki}} = L_s \cdot \frac{I_{s_{ki}}}{\sqrt{2}} \pm j \cdot L_{sc} \cdot \frac{I_{s_{ki}}}{\sqrt{2}}; \quad \bar{I}_{s_{ki}} = \frac{I_{s_{ki}}}{\sqrt{2}} \cdot (1 \pm j) \quad (7.38)$$

with a positive sign (+) for motoring and a negative sign (−) for generating.
The stator voltage is simply

$$\begin{aligned} \bar{V}_{s_{ki}} &= R_s \cdot \bar{I}_{s_{ki}} + j \cdot \omega_1 \cdot \bar{\Psi}_{s_{ki}} \\ &= R_s \cdot \frac{I_{s_{ki}}}{\sqrt{2}} \mp \omega_1 \cdot L_{sc} \cdot \frac{I_{s_{ki}}}{\sqrt{2}} + j \cdot \left(\pm R_s \cdot \frac{I_{s_{ki}}}{\sqrt{2}} + \omega_1 \cdot L_s \cdot \frac{I_{s_{ki}}}{\sqrt{2}} \right) \end{aligned} \quad (7.39)$$

$$\begin{aligned} V_{s_{ki}} &= \frac{I_{s_{ki}}}{\sqrt{2}} \sqrt{(R_s \mp \omega_{1b} \cdot L_{sc})^2 + (R_s \mp \omega_{1b} \cdot L_s)^2} \\ \omega_{rb} &= \mp \frac{Rr}{Lr} + \omega_{1b} \end{aligned} \quad (7.40)$$

The maximum stator voltage vector fundamental $V_{s_{ki}}$ is dependent mainly on the battery voltage V_{dc} and, to a smaller degree, on the type of PWM strategy that is applied. In general,

$$V_{s_{ki}} = \frac{4}{\pi} \cdot \frac{V_{dc}}{\sqrt{3}} \cdot k_{PWM} \text{ (peak value per phase)} \quad (7.41)$$

The coefficient k_{PWM} depends on the PWM strategy but in general is in the interval of 0.9 to 0.96. It is important to mention here again that the battery voltage varies with the SOD (SOC) and ambient temperature and load.

There might be a 30 to 50% difference between maximum and minimum value of battery voltage, and to secure the peak torque for the worst case, it might prove to be practical to use a voltage boost bidirectional DC–DC converter in front of a PWM converter instead of oversizing the PWM converter to comply with peak torque current demand at the lowest battery voltage.

Above base speed, the current angle should be increased, sacrificing more current for flux reduction. The best exploitation of voltage corresponds to maximum torque per flux:

$$\begin{aligned} \Psi_s &= \sqrt{(L_s \cdot I_M)^2 + (L_{sc} \cdot I_T)^2} \\ T_e &= \frac{3}{2} \cdot p_1 \cdot (L_s - L_{sc}) \cdot I_M \cdot I_T; \quad I_T = I_M \cdot S \cdot \omega_1 \cdot \frac{L_r}{R_r} \\ \frac{\partial T_e}{\partial I_M} &= 0 \end{aligned} \quad (7.42)$$

Finally, for peak torque at given flux:

$$L_s \cdot I_{Mk\Psi} = L_{sc} \cdot I_{Tk\Psi} = \frac{\Psi_s}{\sqrt{2}} \quad (7.43)$$

and

$$T_e = \frac{3}{2} \cdot p_1 \cdot (L_s - L_{sc}) \cdot \frac{\Psi_s^2}{2 \cdot L_s \cdot L_{sc}}; \quad (\omega_1 S)_{k\Psi} = \frac{R_r}{L_r} \cdot \frac{L_s}{L_{sc}} \quad (7.44)$$

Approximately, for given voltage,

$$\Psi_s \approx \frac{V_s k_V}{\omega_1}; \quad k_V = 0.95 - 0.97 \quad (7.45)$$

The flux level decreases inversely proportional to frequency. The slip frequency $(\omega_1 S)_{k\Psi}$ for maximum torque per flux, though constant, is rather large.

The current angle $(\delta_i)_{k\Psi}$ is, from Equation 7.44,

$$(\delta_i)_{k\Psi} = \tan^{-1} \left(\frac{I_M}{I_T} \right) = \tan^{-1} \left(\frac{L_{sc}}{L_s} \right) \ll \frac{\pi}{4} \quad (7.46)$$

The current power angle (with axis q) for maximum torque per flux is much smaller than 45° in this case. The maximum torque per flux condition (and $\delta_{ik\Psi}$) should be used in the design to define *the maximum speed* for constant power, either motoring or generating.

Beyond this frequency (speed) — $\omega_{1\max} = \omega_{r\max} \pm (\omega_1)$ — producing constant power is not feasible, but some power is available, though at higher losses per Newton meter of torque.

It is, thus, clear that the current power angle δ_i , if controlled, should be $\delta_i = (\delta_i)_{ki} \approx \frac{\pi}{4}$ up to base speed and continuously decreasing above this speed down to $\delta_i = (\delta_i)_{ki} = \tan^{-1}(L_{sc}/L_r)$ for maximum constant power speed. Also, the slip frequency increases from $(\omega_1)_{ki} = \frac{R_r}{L_r}$ to $(\omega_1)_{k\Psi} = \frac{R_r}{L_r} \cdot \frac{L_s}{L_{sc}}$ (Figure 7.12).

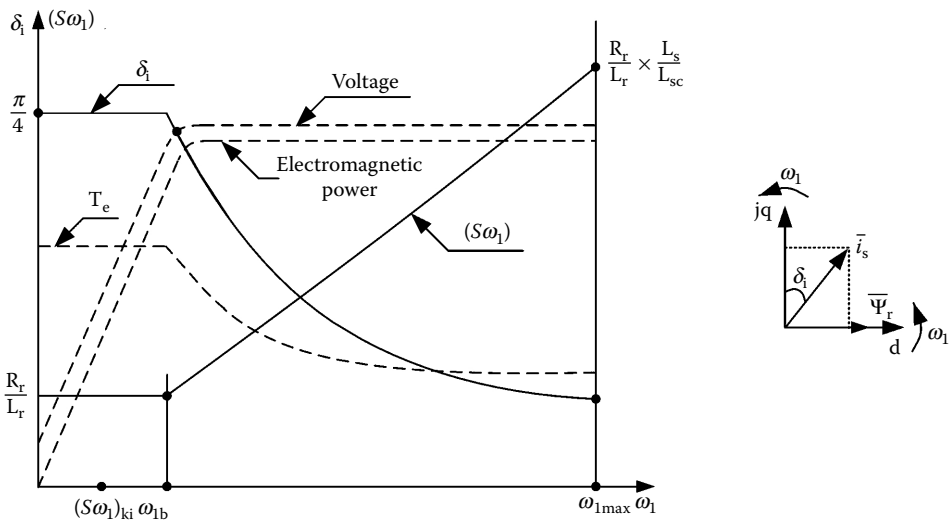


FIGURE 7.12 Current power angle δ_i and slip frequency $S\omega_1$ vs. stator frequency ω_1 for constant torque and then constant power and voltage.

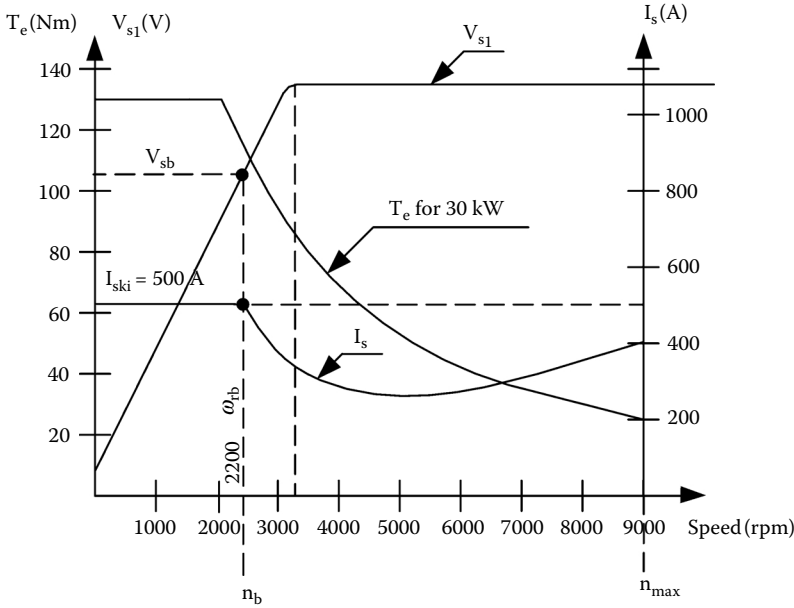


FIGURE 7.13 Voltage and current vs. speed envelopes for constant torque (up to 2000 rpm) and constant power (30 kW) from 2200 rpm to 9000 rpm.

As a conclusion of the above rationale, we mention that once the peak torque T_{ek} , base frequency ω_{lb} , and maximum frequency ω_{lmax} (for constant power) are fixed, the main machine design options are made.

For constant electromagnetic power P_{elm} :

$$P_{elm} = \frac{T_e \omega_l}{P_l} \quad (7.47)$$

and constant voltage, from ω_{lb} to ω_{lmax} , the reference flux and torque may be calculated easily from Equation 7.44 and Equation 7.47. Then, from Equation 7.42, the required values of stator current orthogonal components I_M and I_T (and δ_i) and $S\omega_l$ may be calculated.

The stator current for each frequency is then calculated from Equation 7.42.

Such typical calculations for 30 kW constant power are shown in Figure 7.13 [9].

It should be noticed that, though the constant torque is provided up to 2000 rpm, the voltage V_s is lower than V_{ski} (the maximum value). This means oversizing the converter (in current), but it provided the requested 4:1 constant power speed range without winding reconfiguration.

Once the above theoretical tool is in place, the machine magnetization curve and parameters are known, other steady-state characteristics, such as machine losses and current power angle vs. speed may be determined.

7.5 Vector Control of ISA

A generic vector control system for ISA is as shown in Figure 7.14. It has the following main components:

- The reference rotor flux Ψ_r^* calculator (for handling both motoring and generating)
- The reference torque T_e^* calculator (for motoring [$T_e^* > 0$] and generating [$T_e^* < 0$])

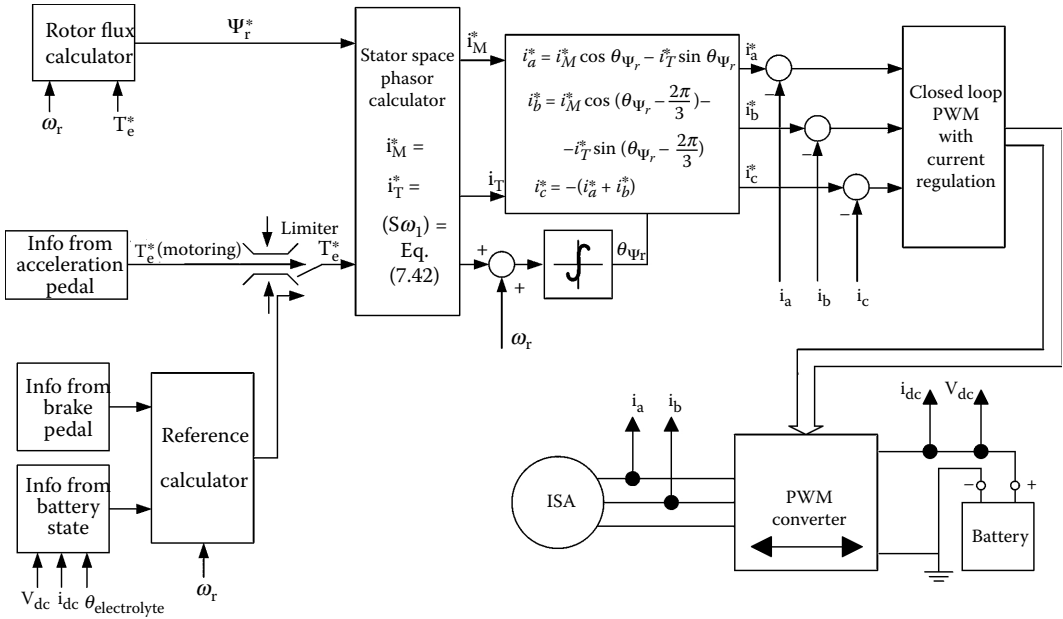


FIGURE 7.14 Generic (indirect alternating current [AC]) vector control system for induction starter/alternators (ISAs).

- The stator space-vector current components calculator, based on Equation 7.42, adapted to constant torque and constant power conditions as required for motoring and generating below and above base speed, for battery recharging, and for regenerative break
- The vector rotator which transforms the currents vector from rotor flux to stator coordinates
- The closed-loop PWM system based on alternating current (AC) regulators

It should be noticed that the state of the acceleration and brake pedals and of the battery and the speed have to be considered in the reference flux and torque calculators, in order to harmonize the driver's motion expectations with energy conversion optimal flow on-board.

In a direct current vector system, the rotor flux position θ_{Ψ_r} , rotor flux, and speed may all be estimated [14], and thus, a motion-sensorless system may be built. The vehicle has to start firmly, even from a stop on a slope; thus, firm and fast torque responses are required from zero speed. Only for cruise control is an external speed regulator added.

So, all sensorless systems have to provide safe estimation of θ_{Ψ_r} and Ψ_r at zero speed. This is how signal injection solutions became so important for sensorless ISA control [16].

The signal injection observer of θ_{Ψ_r} , Ψ_r and ω_r has to be dropped above a certain speed due to large losses in the machine, inverter voltage usage reduction, and hardware and software time and costs. The transition between the two observers has to be smooth [16].

The above indirect current vector control scheme [14] has the merit that it works from zero speed in the torque mode, but adaptation for rotor resistance R_r and for magnetic saturation have to be added.

If a speed sensor is available on an EHV, the indirect current vector control with rotor resistance adaptation and saturation consideration constitutes a practical solution for the application.

7.6 DTFC of ISA

DTFC provides closed-loop control of flux and torque that directly triggers the adequate voltage vector in the converter. To reduce torque and current ripple, a regular sequence of neighboring voltage vectors with a certain timing is needed. For the basic DTFC, please see Chapter 6. A general scheme for DTFC for ISA is shown in Figure 7.15.

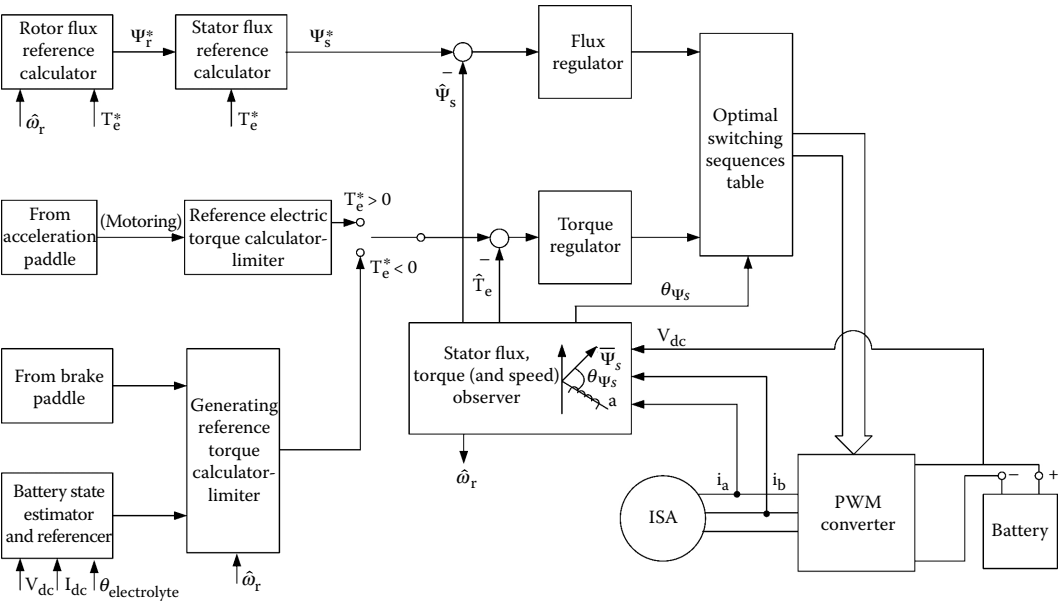


FIGURE 7.15 Direct torque and flux control (DTFC) of induction starter/alternator (ISA).

For DTFC, the AC (or DC) current regulators are replaced with DC stator flux and torque regulators. Also, a state observer that calculates stator flux amplitude Ψ_s , and position angle θ_{Ψ_s} (in stator coordinates) and then calculates the torque, is required. In motion sensorless configurations, a speed observer is also needed.

As speed control is not imperative at very low speed, sensorless DTFC without signal injection was proven to produce fast and safe torque response at zero speed [17] (Figure 7.16a and Figure 7.16b). Variable structure (sliding mode) control was used both in state observer and in the torque and speed regulators.

The sliding mode flux observer [17], shown in Figure 7.17, combines the voltage model in stator coordinates and the current model in rotor flux coordinates to eliminate the speed estimation interference (errors) in the observers. The two models are connected through a sliding mode current error corrector

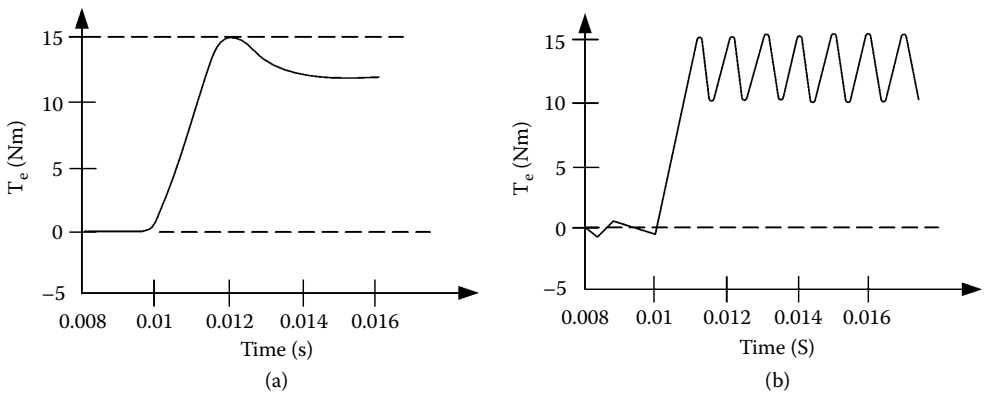


FIGURE 7.16 Step torque response of motion-sensorless direct torque and flux control (DTFC) at zero speed without signal injection: with (a) sliding mode control and space-vector modulation and (b) classical DTFC.

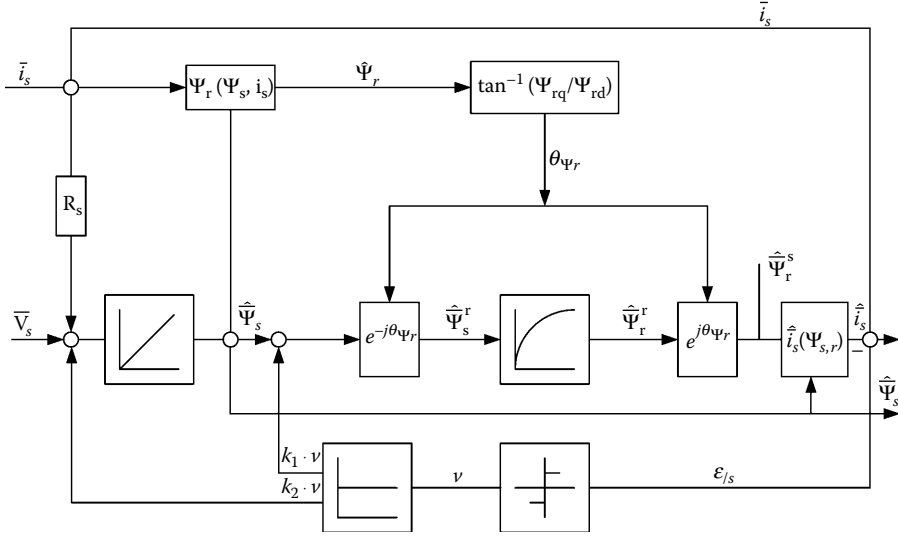


FIGURE 7.17 Sliding mode flux observer.

(the current is also estimated, not only measured). This way, the current model prevails at low speed and the voltage model at high speeds.

A stator resistance corrector is needed for precision control at very low speeds. The rotor resistance is considered proportional to that of the stator:

$$\hat{R}_r = R_{so} - k_{RS} \cdot (\hat{\Psi}_{rd} \cdot (i_{sq} - \hat{i}_{sq}) - \hat{\Psi}_{rq} \cdot (i_{sd} - \hat{i}_{sd})) \quad (7.48)$$

$$\hat{R}_r = k_{sr} \cdot \hat{R}_s \cdot \frac{R_{ro}}{R_{so}} \quad (7.49)$$

i_s is the measured and $\hat{i}_s$ the estimated current vector (Figure 7.17).

The stator resistance adaptation greatly reduces the flux estimation errors [17].

The torque calculator is straightforward:

$$\tilde{T}_e = \frac{3}{2} \cdot p_1 \cdot \text{Real}(j \hat{\Psi}_s \cdot \bar{i}_s) \quad (7.50)$$

A few solutions for the speed observer may be applied for the scope [14], but as speed control of ISA is not required at low speeds, a standard solution may be appropriate for the case:

$$\hat{\omega}_r = \hat{\omega}_{\Psi_r} - (S\hat{\omega}_1) \quad (7.51)$$

$$\hat{\omega}_{\Psi_r} = \frac{\text{Real}\left(\hat{\Psi}_r \cdot j \frac{d\hat{\Psi}_r}{dt}\right)}{|\hat{\Psi}_r|^2} \quad (7.52)$$

$$(S\hat{\omega}_1) = \frac{2}{3} \frac{\hat{R}_r}{p_1} \frac{\tilde{T}_e}{|\hat{\Psi}_r|^2} \quad (7.53)$$

The rotor speed comes as the difference between the rotor flux vector speed and the slip speed. An exact value of rotor resistance $\hat{R}_r$ is required for good precision. This justifies the rotor resistance adaptation, as in the flux estimator. The rotor resistance interferes only in the current model (that is, at low speeds).

The control hardware and software do not change with speed.

Note that while both vector control and DTFC are capable of similar dynamic and steady-state performance, DTFC seems slightly superior when torque control is needed and motion-sensorless control is preferred.

7.7 ISA Design Issues for Variable Speed

There are a few design peculiarities to ISA design for variable speed. We already mentioned a few in previous paragraphs. Here we present them in a more systematic manner.

7.7.1 Power and Voltage Derating

There is a rich body of knowledge on induction machine design (mostly for the motor operation mode) for constant voltage and frequency [13].

The Epson's constant C_o (W/m³), as defined by past experience, is an available starting point in most standard designs. To apply it to ISA, we first have to reduce C_o to account for additional core and winding time harmonics losses. Then we have to increase it for peak torque requirements in constraint volume.

With today's insulated gate bipolar transistor (IGBT) PWM voltage source converters for DC battery voltage above 200 V, and power MOSFETs for less than 200 V_{dc} batteries, the converter derating of ISA is $P_{derat} = 0.08$ to 0.12.

The Epson's constant is of little value for ISA designed for speeds above 6000 rpm, belt-driven, as little experience was gained in the subject.

Voltage derating is due to PWM converter voltage drops. It amounts to 0.04 to 0.06 P.U. for above 200 V_{dc} batteries voltage, but it may go well above these values for 42 V_{dc} batteries.

In designs that are tightly volume constrained, such as ISA, it may be more appropriate to use as a design starting point the specific tangential peak rotor force density (shear stress) f_t (N/cm²).

Peak values from four to almost 12 N/cm² may be achieved with current densities ranging from 10 to 40 A/mm². Naturally, forced cooling is generally necessary for ISAs.

As the battery voltage varies from V_{dcmín} to V_{dcmáx} by 30% or more, the design may be appropriate for average rated V_{dc} with verifications on performance for minimum battery voltage.

7.7.2 Increasing Efficiency

Increasing efficiency is important to ISA to save energy on board vehicles. Volume constraint is contradictory to high efficiency, and trade-offs are required.

While volume constraints lead inevitably to increased fundamental winding and core losses, there are ways to reduce the additional (strayload) losses due to space and time harmonics.

Space harmonics are mainly due to stator and rotor slotting and magnetic saturation, but time harmonics are mainly due to PWM converter supply (Reference [13], Chapter 11 on losses).

A few suggestions are presented here:

- Adopt a large number q_1 of slots/pole/phase, if possible, in order to increase the order of the first space slot harmonic ($6q_1 \pm 1$).
- Compare thoroughly designs with different pole counts for given specifications.
- In long stack designs, use insulated or at least noninsulated rotor cage bars with high bar-contact resistance in skewed rotors to reduce interbar current losses.
- Use $0.8 < N_r/N_s < 1$ (N_s , N_r stator and rotor slot count) in order to reduce the differential leakage inductance of the first slot harmonics pair $6q_1 \pm 1$ and, thus, reduce interbar rotor current. Skewing may not be needed in this case, provided the parasitic torques are within limits.

- Skewing is necessary for $q_1 = 1, 2$; even a fractionary winding, with all coils in series, and free of subharmonics ($q = 1\frac{1}{2}$), may be tried in order to reduce strayload losses.
- Thin (less than 0.5 mm thick) laminations are to be used in ISA design where the fundamental frequency is above 300 Hz, to reduce all core losses.
- Use chorded coils to reduce end turns (and losses) and the first phase belt harmonics (5,7) parasitic asynchronous torque.
- Carefully increase the airgap to reduce additional surface losses, but check on power factor reduction (peak current increases).
- Re-turn rotor surface to prevent lamination short-circuits that may produce notably higher rotor surface additional core losses.
- Use sharp stamping tools and special thermal lamination treatment to reduce fundamental frequency core losses [18] above 300 Hz.
- Use copper rotor bars whenever possible to reduce the rotor cage losses and rotor slot size.

7.7.3 Increasing the Breakdown Torque

Large breakdown torque by design is required when a more than 2:1 constant power speed range is desired. This is the case of ISA, where ratios ω_{max}/ω_{rb} above 4:1 are typical and up to 10(12):1 would be desirable. Breakdown to rated torque T_{ek}/T_{es} ratios in induction machines is in the 2.5 to 3.5 range. The natural constant power ideal speed range coincides with the T_{ek}/T_{eb} ratio. When

$$\frac{\omega_{rmax}}{\omega_{rb}} > \frac{T_{ek}}{T_{eb}} \quad (7.54)$$

machine oversizing and other means are required.

Still, a high T_{ek}/T_{eb} ratio is desirable without compromising too much efficiency and power factor. As the breakdown torque T_{ek} may be approximated to ($R_s = 0$),

$$T_{ek} \approx \frac{3}{2} \cdot P_1 \cdot \left(\frac{V_s}{\omega_1} \right)^2 \cdot \frac{1}{2L_{sc}} \quad (7.55)$$

reducing the short-circuit inductance L_{sc} of ISA is the key to higher breakdown torque.

The two components — stator and rotor — of short-circuit (leakage) inductance are as follows (Reference [13], Chapter 6):

$$L_{sl} = 2\mu_o l_{stack} \cdot \frac{w_1^2}{p_1 \cdot q_1} \cdot (\lambda_{ss} + \lambda_{zs} + \lambda_{ds} + \lambda_{end}) \quad (7.56)$$

$$L_{rl} = 4ml_{stack} \cdot \left(\frac{w_1 \cdot k_{w1}}{N_r} \right)^2 \cdot 2\mu_o (\lambda_b + \lambda_{er} + \lambda_{zr} + \lambda_{dr} + \lambda_{skew}) \quad (7.57)$$

with

- λ_{ss}, λ_b the stator (rotor) slot permeance coefficients
- $\lambda_{zs}, \lambda_{zr}$ the stator (rotor) zigzag permeance coefficients
- $\lambda_{ds}, \lambda_{dr}$ the stator (rotor) differential leakage coefficients
- $\lambda_{end}, \lambda_{er}$ the stator (rotor) end connection (ring) leakage coefficients
- λ_{skew} the rotor skewing leakage coefficient
- w_1 the turns per phase
- p_1 the pole pairs

With so many terms in Equation 7.56 and Equation 7.57, an easy way to reduce L_{sc} is not self-evident. However, the main parameters that influence L_{sc} are as follows:

- Pole count: $2p_1$
- Stack length/pole pitch ratio: l_{stack}/τ
- Slot/tooth width ratio: $b_{ss,sr}/b_{t_s,t_r}$
- Stator slots/pole/phase: q_1
- Rotor slots/pole pair: N_r/p_1
- Slot opening per airgap: $w_{os,or}/g$
- Stator and rotor slot aspect ratio: $h_{ss,sr}/b_{ss,sr}$
- Air flux density: B_{g1}
- Stator (rotor) peak torque current density: j_{sk}, j_{rk}

Parameter sensitivity analyses showed that $2p_1 = 4, 6$ are best suited for speeds above 6000 rpm and $2p_1 = 8, 10, 12$ for speeds below 6000 rpm when wide constant power speed range conjugated with high peak torque at standstill are needed in volume constraint designs such as ISAs. The differential leakage tends to decrease with increased q_1 slots/pole/phase and decreased slot aspect ratios $h_{ss,sr}/b_{ss,sr} < 3.5$ to 4. However, this tends to increase the machine volume or oversaturate the core.

Longer core stacks l_{stack} may allow for a smaller stator bore diameter and, thus, shorter end connection in the stator windings. Slightly less end-connection leakage inductance is obtained this way. Only in high-speed ISAs with four poles does such a strategy seem practical.

Skewing, if avoided, as explained in the previous paragraph, may contribute to L_{rl} (and, thus, L_{sc}) reduction. Special stator windings with four layers were proven to reduce, in a few cases, by 30%, the stator leakage inductance [19]. However, this occurred at the price of added winding manufacturing costs.

7.7.4 Additional Measures for Wide Constant Power Range

Beyond the natural constant power speed range (T_{ek}/T_{eb} ratio: 2.5 to 3.5) illustrated in Figure 7.18a, machine oversizing is required, as the base torque T_{eb} has to go below rated (continuous) torque T_{er} (Figure 7.18b).

When the constant power speed range is larger than 4:1, machine oversizing will not do it alone, and winding reconfiguration or pole count changing is required.

7.7.4.1 Winding Reconfiguration

Consider a wide constant power speed range $c_\omega = \omega_{rmax}/\omega_{rb} > 4$ and allow for a torque reserve at all speeds above rated (base) torque requirements (Figure 7.19):

$$\frac{T_{ek}}{T_{eb}} = c_{bT} > 1$$

$$\frac{T_{eMK}}{T_{eM}} = c_{MT} > 1 \quad (7.58)$$

$$\frac{T_{eM}}{T_{eb}} \approx \frac{\omega_{rb}}{\omega_{rmax}} = \frac{1}{c_\omega} \ll 1$$

As in this case $T_{ek}/T_{eb} < \omega_{rmax}/\omega_{rb}$, the phase voltage required to provide the wide constant power range should not reach its maximum value at base speed but somewhere below it.

We may consider T_{ek} and T_{eMK} as breakdown torques; thus,

$$T_{ek} \approx \frac{3}{2} \cdot p_1 \cdot \left(\frac{V_b}{\omega_b} \right)^2 \cdot \frac{1}{2L_{sc}} \quad T_{eMK} = \frac{3}{2} \cdot p_1 \cdot \left(\frac{V_{max}}{\omega_{max}} \right)^2 \cdot \frac{1}{2L_{sc}} \quad (7.59)$$

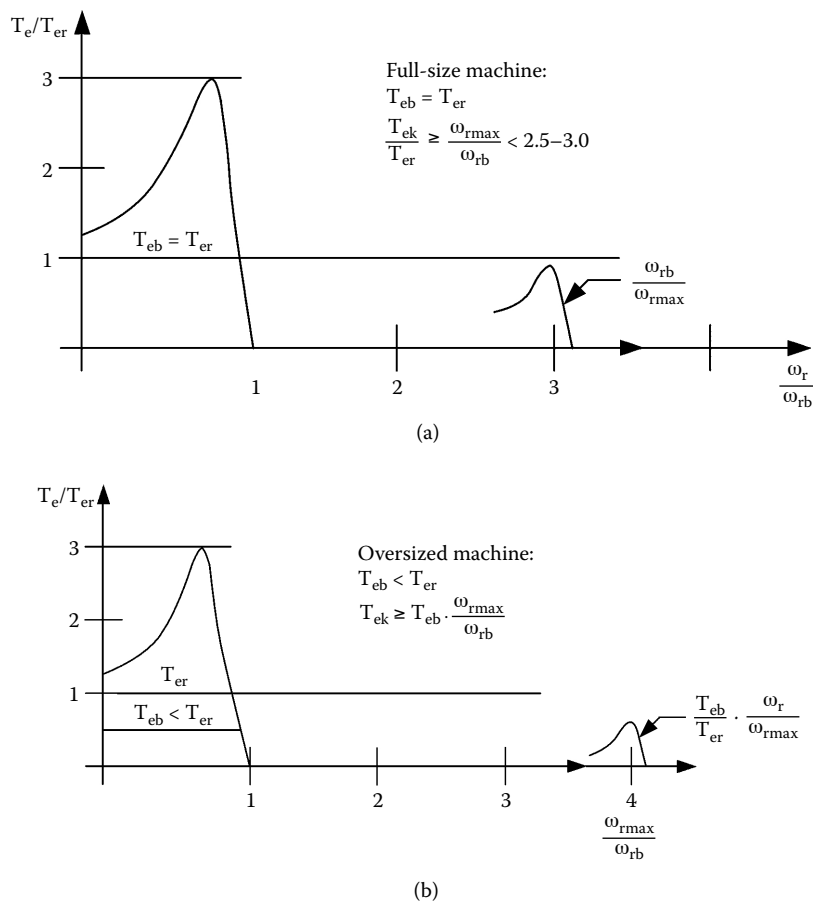


FIGURE 7.18 Natural constant power speed range: (a) full-size machine and (b) oversized machine.

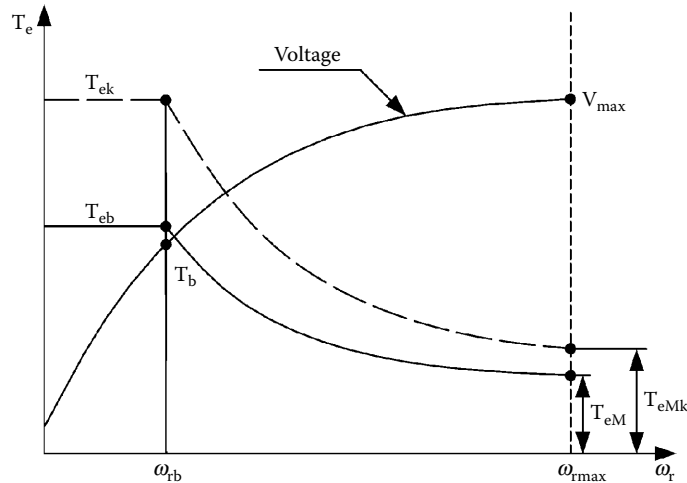


FIGURE 7.19 Phase voltage increasing for extended constant power speed range.

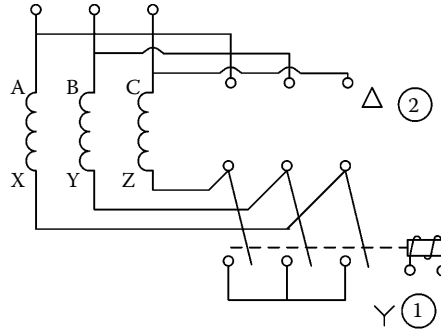


FIGURE 7.20 Star (Y) to delta (Δ) winding changeover.

Consequently,

$$\left(\frac{V_{max}}{V_b} \right)^2 = \left(\frac{\omega_{max}}{\omega_b} \right)^2 \cdot \frac{T_{eMK}}{T_{ek}} = \frac{c_{\omega} \cdot c_{MT}}{c_{bT}} \quad (7.60)$$

For $\omega_{max}/\omega_b = 4$, $c_{bT} = 3$, $c_{MT} = 1.5$, $V_{max}/V_b = 1.41$.

Consequently, a 30% reduction in the voltage has to be allowed at base speed. Accordingly, the same increase in the phase current is to be allowed for. Consequently, the PWM converter has to be oversized by 1.41 times in current. The machine is to be oversized in this situation only if it is not capable of a 4:1 peak torque/rated torque ratio, which is the case in most situations. Realistic values of the above torques may be calculated by making use of induction machine standard equivalent circuits. A further increase in the constant power speed range requires winding changeover. The obvious choice is from star to delta connection (Figure 7.20).

Only a dual action power electromagnetic switch is required to perform the star to delta switch. During this changeover, the PWM converter control should be inhibited. If the time required for the changeover is less than 0.1 sec, it will hardly be felt by the EHV power system.

When the winding connection switches from star to delta, it is as if the phase voltage was increased $\sqrt{3}/1$ times.

Consequently, the constant power speed range, proportional to voltage squared, because the torque is proportional to voltage squared, is increased three times:

$$\frac{\omega_{max_delta}}{\omega_{max_star}} = \left[\frac{(V_{phase})_{delta}}{(V_{phase})_{star}} \right]^2 = \frac{3}{1} \quad (7.61)$$

With this 3:1 ratio, the method is efficient for the scope and is in industrial operation in some spindle drives. It should work as well for EHV.

Winding tapping would bring a similar increase in voltage, but only part of the machine winding would remain at work at high speeds. So, the overheating of the machine due to increased core losses has to be attended for properly.

So, it seems that the star to delta switching changeover is the best option.

Note that inverter pole changing in the ratio of 2:1 with a Dahlander winding may produce an almost 2:1 constant power speed range by using two twin PWM converters for the two half-windings. For one pole count, the currents in the two converters are in phase, and for the other they are opposite. The transition may be smooth [20], but still at the price of two converters, even designed at half the rating of a single one, because full battery voltage is available for both of them.

Note that for the smaller pole count, the winding factor goes down to 0.6 to 0.7 from 0.86 to 0.933, and its end connection leakage inductance and resistance are larger than they should be in a dedicated winding. This fact leads to a notable reduction of the actual constant power speed range from 2 to probably less than 1.5.

At a higher ratio (3:1 for example), a pole count changing winding with high performance is not apparently available in an easy-to-manufacture-and-connect winding, so this latter approach seems to be less practical for industrial purposes.

Example 7.1

Consider a belt-driven ISA. The belt transmission ratio is 3:1, and the ISA has to deliver the peak torque of 140 Nm up to 500 rpm and the constant peak power (7.326 kW) up to 6000 rpm (engine speed). Only $7.326/3 = 2.432$ kW is required from ISA for continuous generating from 500 to 6000 rpm: the breakdown to rated torque ratio of the machine $c_{bT} = T_{bk} / T_{eb} = 3.0$.

It is requested that we calculate the PWM converter current oversizing and machine peak torque oversizing to perform the $c_{bT} = \omega_{\max} / \omega_{eb} = 6000/500 = 12/1$ constant power speed range.

Solution

The star to delta switching can produce a theoretical 3/1 speed range increase. So, it will be placed at $c_{bT}/3 = 4.0$ base speed. That is, the star to delta switching will take place at 4.0 times base speed (Figure 7.21).

Consider a zero reserve of torque at maximum engine speed $c_{MT} = 1$. Also, at base speed, $c_{bt} = 3.0$ and $\omega_{\max_{star}} / \omega_b = 4/1$.

Then, from Equation 7.60, applied at $4\omega_b$ for delta connection, and at ω_b for star connection, the respective voltage ratios are calculated:

$$\left(\frac{V_{M_{star/delta}}}{V_{rated}} \right)_B = \sqrt{\frac{c_{bT}}{c_{\omega_{star/delta}} \cdot c_{MT}}} = \sqrt{\frac{3}{3 \cdot 1}} = 1$$

$$\left(\frac{V_{rated}}{V_b} \right)_A = \left(\frac{V_{rated}}{V_{M_{star/delta}}} \right) \cdot \sqrt{\frac{\omega_b}{\omega_{\max_{star}}} \cdot \frac{c_{bT}}{c_{MT}}} = 1 \cdot \sqrt{\frac{3}{4 \cdot 1}} = 0.866$$

For the same power factor, the design current I'_b at base speed (point A in Figure 7.21) is increased with respect to that of base (max) voltage, I_b , by

$$\frac{I'_b}{I_b} = \frac{V_b}{V_{rated}} = \frac{1}{0.866} = 1.1547$$

The same ratio is valid for the peak torque current ratios in the two cases.

The converter oversizing is equal to current overloading, which is only 15.47% in our case. However, the machine oversizing in torque is equal to the constant power speed ratio for star connection (4/1) to peak/rated torque ratio (3/1); that is, $4/3 = 1.33$, or 33%.

As the machine costs are notably lower than the PWM converter costs for given kilovoltampere (especially for battery voltages $V_{dc} < 200$ V), the above division of oversizings seems to be adequate.

As in our case, for generator mode, three times less than peak power is required, there should be no problem in obtaining it with proper control. Even star connection would do for generating in our

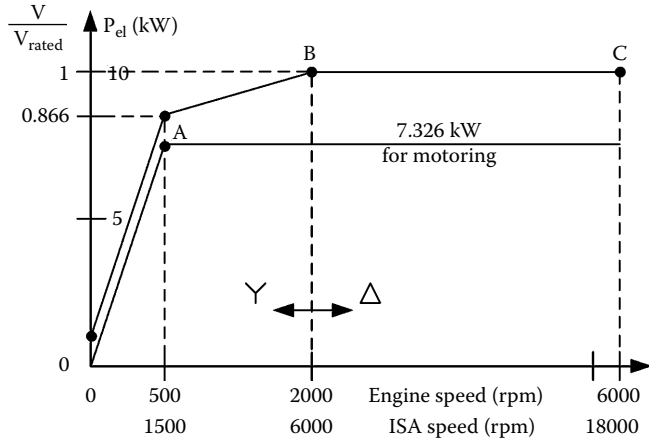


FIGURE 7.21 Voltage for wide speed constant peak power motoring of an induction starter/alternator (ISA) with Y and Δ winding switching.

case, but performing star to delta and delta to star switching at a given speed may prove more practical to implement in the control, in order to limit the wearing-off of the electromagnetic power switches.

With the exception of generating limitations at high speed, attributed here in part to control insufficiencies, very good performance in torque/speed envelopes was proven in Reference [6] and Reference [15] for full-size ISAs for mild and, respectively, full hybrid electric cars.

Still, the rather high rotor winding losses at low speed for peak torque constitute a severe obstacle to ISA industrialization.

In Reference [22], an urban electric vehicle with ISA is presented with test results for 14 kW constant power at basically a 3:1 speed range.

A design comparison between two electric vehicles with ISA, one for in-wheel placement (400 rpm, rated speed) and one with a 10/1 gear ratio, at 5 kW, was performed considering the losses in the ISA and in the converter for a 3:1 constant power speed range, for urban- and country-drive standard European cycles.

With stator flux adapted to torque, a notable energy savings for in-town drive is obtained. The geared ISA is shown to peak slightly superior in terms of losses (or range) at 16.3 kg weight (ISA plus gearbox) vs. 40 kg for the in-wheel configuration [23].

Yet another such comparison design of direct-driven and transmission-driven ISA for a mild hybrid vehicle (at $42 V_{dc}$) for 6.5 kW generating from 1000 to 6500 rpm engine speed and 350 Nm crankshaft starting torque is given in Reference [24]. Theory and some test results are offered. Winding changeover is used, as a 6.5/1 constant power speed range is required.

The epicyclic gear is operational only in the motor mode. This way, the ISA peak torque is reduced. A 12-pole ISA is adopted to reduce the slip frequency and losses. The power factor increases, and thus, the peak current is reduced. The battery requirements and PWM converter rating are also reduced.

For the gearless solution, the star to delta switch of the winding is used to reach the large constant power speed range in an 18-pole ISA configuration.

About 6.5 N/cm² shear rotor stress for the peak torque is adopted in designs.

Both solutions have their definite merits.

Note that a comprehensive design methodology of an ISA, complete for specifications, should also consider the PWM converter and battery losses and costs in order to provide an industrial design. For simplified PWM loss modeling, with and without DC voltage booster, see Reference [9].

Blending the standard induction machine design body of analytical knowledge with finite element method (FEM) verifications (see Reference [13] for details) for the ISA special operational conditions, should lead to such an industrial multiobjective optimization design of the ISA system. Such a grueling task is beyond our space and scope here.

7.8 Summary

- “Electrification” of road vehicles is under aggressive development today with some dynamic industrial markets in place.
- The degree of electrification is defined by the electric fraction %E, the ratio between peak electric propulsion power and peak total power of the vehicle.
- With %E < (70 to 80)%, the vehicles are called hybrids; with %E = 100%, they are called electric vehicles.
- High peak starting torque and wide constraint power speed range (up to 12:1) are required in EHVs under constraint volume and costs, but with low total system losses, to cut fuel consumption and increase battery life.
- Given its ruggedness and low cost, the induction machine seems a natural choice for starter/alternators on EHVs. They are called here induction starter/alternators (ISAs).
- In mild hybrids, the starter/alternator is to be designed for low battery voltage: the 42 V_{dc} bus, in general. It uses a PWM power converter interface to deliver controlled torque for the entire speed range. It is also limited in power to (in general) a maximum of 6 kW.
- Full hybrids have ISAs that deliver above 6 kW of power for both motoring and generating and require high voltage battery packs (180 V_{dc} and more). ISAs assist the driving substantially at low speeds (up to 2000 rpm engine speed) and still notably above this speed. Powers of 30 to 50 kW are typical for full hybrid cars and above 75 kW for hybrid city buses. Electric vehicles with batteries or fuel cells or inertial batteries (flywheels) require even more power, as electric propulsion acts alone.
- The battery voltage is a very important design factor for ISA, and its modeling is crucial to determine the battery receptivity during its normal recharging (at high speed) or during regenerative braking by ISA.
- Battery capacity Q, discharge rate Q/h, SOC, SOD, and DOD, are among the main parameters in battery modeling. Linear models, with variable emf (e_v) and internal resistance R_i are typical.
- The number of charge–discharge cycles, the specific energy (Wh/kg) and specific peak power (W/kg) and efficiency characterize the battery performance for EHVs.
- An actual lead-acid battery is characterized by about 50 Wh/kg, 400 W/kg, 80% efficiency, 500 to 1000 cycles at 150 \$/kWh. Newer batteries such as lithium polymer batteries are credited in tests with 150 to 200 Wh/kg, 350 W/kg, 1000 cycles at 150 \$/kWh. Further developments are needed to dethrone the lead-acid batteries, with valve-regulated electrolyte, from road vehicles, because they are an established technology.
- Super-high-speed flywheels (in vacuum) with magnetic suspension and special rotor composite materials have been tested up to 1 km/sec peripheral speed, when the 50 Wh/kg of lead acid battery is surpassed at 2000 W/kg and 2 to 3 min “recharging.” The recharging of the flywheels is performed through the generator/motor on its shaft and its PWM power interface converter.
- Given the challenging torque, efficiency, power factor, speed range, under constraint volume, and costs of the ISA system, quite a few design particularities characterize them.

- As in any variable speed system, the performance and costs prevail. This way, the fundamental frequency in the ISA is limited, with silicon lamination core, to 500 to 600 Hz, to also reduce the PWM converter switching losses.
- Depending on the usage (or not) of the transmission between the ISA and the ICE engine (in hybrid vehicles) and between ISA and wheels in electric vehicles, the maximum speed of ISA varies from 2000 rpm in city buses to 18,000 rpm in the mild hybrids with belted ISAs.
- Consequently, ISAs placed on the ICE shaft are designed with 8, 10, or 12 poles, while the belted ones have four or six poles.
- Skin effect is not desirable, and this restricts the shape of the rotor slot.
- The high peak temperature in the rotor (240°C or more) tends to suggest using copper bars in the rotor cage.
- Advanced magnetic saturation is needed to reach high peak torque density at low speeds and for start. Uniform oversaturation of stator and rotor teeth and yokes reduces the airgap flux density space harmonics. So, provided the magnetization curve of ISA is known from analytical or FEM models, the equivalent circuit and the space-phasor model may be used in the ISA design, performance, computation, and control.
- Constant rotor flux (rotor flux orientation) — FOC — allows for fastest transient torque of ISA in rotor flux coordinates.
- For constant rotor flux amplitude in synchronous coordinates ($\omega_b = \omega_i$), the ISA acts as a high saliency reluctance synchronous machine; the virtual saliency is “produced” by the rotor current vector, which is at 90° with respect to the rotor flux vector. The d axis inductance is the no-load inductance L_s , and the q inductance is the short-circuit inductance L_{sc} . The d axis current is the rotor flux current i_M , and the torque current, along axis q , is i_T .
- For peak torque production, the maximum torque/current criterion is used. Ideally, $i_{MK} = i_{TK} = i_{SK}/\sqrt{2}$, but due to variable (heavy) magnetic saturation, $i_{MK} > i_{TK}$. This should be embedded in the FOC below base speed.
- For maximum speed constant power production, the maximum torque per available stator flux criterion is used when $i_M = i_T \cdot L_{sc}/L_s$.
- The d - q current angle $\delta_i = \tan^{-1}(i_M/i_T)$ varies from large values at low speed to low values at high speeds.
- Indirect vector control works naturally for zero speed, with magnetic saturation and rotor resistance adaptation. In the presence of a speed sensor, it seems to be a practical control strategy for ISA.
- The rotor flux and torque reference calculators mitigate the driver motion expectations and energy management on board (with battery state as crucial).
- Still, the online computation effort is high.
- DTFC — used commercially by some manufacturers of electric drives — is also feasible for ISA. DTFC uses the rotor flux and torque reference calculators, but it calculates the required stator flux reference and then close-loop regulates the stator flux and torque to trigger a preset sequence in the PWM converter.
- Instead of vector rotation and sophisticated PWM modulation with current regulators (eventually with emf compensation), typical to FOC, DTFC needs a flux observer and a torque calculator based on measured voltage and current.
- As the speed estimation may be obtained out of the flux observer, a motion-sensorless control system is inherent to DTFC for ISA, where torque control is predominantly used. Only during cruise control, is slow speed regulation performed above a certain speed.
- Fast and safe torque response at zero speed, in motion-sensorless DTFC, was reported without signal injection. Operation at zero speed, sensorless, with signal injection is implicit, but injection has to be abandoned above a small speed, for a different strategy, to cover the rest of the large speed range.

- A few design aspects of ISA require special attention to produce comparative results:
 - Power and voltage derating due to PWM converter
 - Design battery DC voltage
 - Ways to decrease strayload losses
 - Increasing the breakdown torque (by decreasing L_{sc})
 - Winding reconfiguration for constant power speed range extension: a 3:1 extension is obtained by star to delta switching; up to 12:1 total constant power speed range could be obtained by some motor and converter oversizing
- In principle, ISA is capable of complying with the challenging EHV requirements, except for the increase in current and losses at high speeds. In terms of PWM converter peak kilovolt-ampere (rating), the ISA performs moderately, though notably better than the surface-permanent magnet rotor brushless machine or the switched reluctance machine.
- A system optimization design, rather than the ISA design optimization, is the way to a competitive solution with ISA.
- This chapter represents a mere introduction to such a daunting task.

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